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Request for grant of a patent

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1. Your reference

   1876701/AM

2. Patent Application Number

   0209372.2 — 24 APR 2002

3. Full name, address and postcode of the or of each applicant *(underline all surnames)*

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   Patents ADP number *(if known)* 7866809002

   If the applicant is a corporate body, give the country/state of its incorporation

   Country: **ENGLAND**
   State:

4. Title of the invention

   RESONATOR FREQUENCY DETECTION METHOD

5. Name of agent

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   Patents ADP number 1826001

6. Priority details

Country	Priority application number	Date of filing

Patents Form 1/77

7. If this application is divided or otherwise derived from an earlier UK application give details

Number of earlier application — Date of filing

8. Is a statement of inventorship and or right to grant of a patent required in support of this request?

YES

9. Enter the number of sheets for any of the following items you are filing with this form.

Continuation sheets of this form	
Description	21
Claim(s)	
Abstract	
Drawing(s)	8 + 8

10. If you are also filing any of the following, state how many against each item.

Priority documents	
Translations of priority documents	
Statement of inventorship and right to grant of a patent *(Patents form 7/77)*	1 + 3 COPIES
Request for preliminary examination and search *(Patents Form 9/77)*	
Request for Substantive Examination *(Patents Form 10/77)*	
Any other documents *(please specify)*	

11. I/We request the grant of a patent on the basis of this application

Signature Beresford & Co
BERESFORD & Co

Date 24 April 2002

12. Name and daytime telephone number of person to contact in the United Kingdom

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Tel: 020 7831 2290

Resonator Frequency Detection Method

This document references the following other patent filings, the contents of which are incorporated by reference herein:

Number	Assignee	Reference
US 4,878,553	Wacom	
GB 95/01095	Synaptics	Spiral I
GB 98/01759	Synaptics	Spiral 3D
WO 99/18653	Synaptics	Commutators for Motors
WO 01/29759	Synaptics	Excitation Drive Technique
WO 00/33244	Synaptics	PDA Spiral System
GB0112332.2	Synaptics	New application 21 May 2001

2 INTRODUCTION

This patent filing is primarily for pen input to computer and communications devices, particularly small, low-cost ones, for example personal digital assistants (PDAs), mobile telephones, Internet browsers and combinations of these. It has particular relevance where those devices are battery powered. It is also relevant to other devices where the resonant characteristics of a system need to be determined, including the applications described in GB 95/01095 and GB 98/02961.

Figure 1 illustrates a typical prior art inductive pen input system. To take a pen position measurement, the Digital Processing and Signal Generating Unit commands a Variable Frequency Generator to drive an excitation circuit 51 in order to excite a magnetic resonator in the pen 41. The pen creates a magnetic field in response to excitation and this is detected as an EMF by sensor coils 31,33,35 and 37. Signals from these coils are synchronously detected by mixers 69, integrated by integrators 71 and then converted to digital form by A to D converter 73. The Digital Processing and Signal Generating Unit uses the resulting digital values to determine the position of the pen and whether the tip is pressed or not.

Figure 2 illustrates key signals that are present in such a system. The Variable Frequency Generator outputs reference signals "In-phase out" and "Quadrature out" at a reference frequency Fex. This signal is used to generate a selected number of cycles of excitation waveform, as shown. In this case the excitation waveform is designed according to the invention described in WO 01/29759. The sensor coil voltages are only mixed and integrated during a period defined by the RX gate. The mixers illustrated here operate by multiplying their input by 1 for mix input =1 and by –1 for mix input =0. When detection is complete, the outputs of the integrators are held and converted to digital form, for example by employing a sloping ADC process.

Typically, an inductive pen sensing system needs to determine the writing force applied to the pen tip. This is typically achieved by designing the pen such that the frequency changes when this pressure is applied, and by designing the position processor such that it can detect this frequency and hence the force on the pen's tip.

GB0112332.2 describes problems with prior art systems, such as the need for a special tuning step before each pen can be use, and discloses means for avoiding this step while retaining the ability to determine whether the pen tip is touching a writing surface. This disclosure builds upon GB0112332.2.

The operation of such a position processor is described below. This description is based on I and Q measurements of the signal from a single sensing coil. That coil may be one of the loop coils described in US 4,878,553 or one of the sin/cos x/y

coils described in GB 98/01759 and WO 00/33244. Alternatively, I and Q information from several coils may be combined. Such combination may be required when information from a single coil is insufficient for unambiguous frequency detection across the whole sensing area. For example, in the case of GB 98/01759 and WO 00/33244 where sensing coils generate zero or low output signals in some positions due to the sinusoidal response pattern.

The results of a pulse echo resonator detection sequence may be modelled by the following set of equations, for frequencies where Fex is sufficiently close to Fpen:

$$r = \frac{Fpen}{Fex}$$

$$A = Ao \cdot e^{-\left(\frac{1-r}{rw}\right)^2}$$

$$P = 2\pi \frac{(1-r)}{rd}$$

$$SI = A \cdot \cos(P + Po)$$

$$SQ = A \cdot \sin(P + Po)$$

Where Fpen is the natural resonant frequency of the pen, Ao is an amplitude term that depends, among other things, on the coupling factor between the pen and sensor coils. Po is a phase offset and SI and SQ are the respective outputs of the I and Q channels connected to a given sensor coil. A defines the amplitude response and P the phase response, both separate functions of frequency. Parameters rw and rd dictate the breadth of the amplitude response and the rate of phase change with frequency respectively. The following description is for Po=0 unless otherwise stated, although with appropriate modification, as known to one skilled in the art, the approaches are equally valid for non-zero values.

Figures 3A and 3B illustrate the response of the system to varying Fpen, with Fex fixed at 106kHz. The x-axis is pen frequency in kHz. Figure 3A is for 8 excitation and detection cycles and figure 3B for 16.

It is possible for a pen sensing system to determine pen frequency by making an estimate of the parameter P above. If the coupling between pen and sensor board is such that it can reverse sign, for example using the coil arrangements described in WO 00/33244, then the following calculation is required to determine an estimate Pe of phase P given the measurements SI and SQ:

$$Pe = a\tan\left(\frac{SQ}{SI}\right)$$

Note that results are returned over a range of π radians, e.g. between $-\pi/2$ and $\pi/2$. If Pe differs from the "ideal" model value P above by a multiple of π, then the phase is said to have wrapped.

If the sign of the coupling factor between pen and sensor board is known, for example if the coils are arrayed as illustrated in US 4,878,553, then phase P may be calculated from:

$$Pe = a\tan 2(SI, SQ)$$

Where atan2 is a function that return the angle between the vector described by its arguments and the vector (1,0).

Note that results are returned over a range of 2π radians, e.g. between $-\pi$ and π. If Pe differs from the "ideal" model value P above by a multiple of 2π, then the phase is said to have wrapped.

In either case, a single measurement of SI and SQ is insufficient to determine Fpen unambiguously unless Fpen is restricted to a range of frequencies such that phase never wraps. As described in GB0112332.2, this may be undesirable since it is preferred to be able to use any pen with such a system without a special tuning step, and this translates to a requirement for a relatively large shift in pen frequency.

It is possible to reduce the phase change for a given frequency change, equivalent to increasing rd in the model above, in order to avoid wrapping. This can be achieved by decreasing the number of excitation or detection cycles (EP or DP respectively). However, as figure 4 illustrates, rw is strongly correlated with rd, so that an increase in rd is always accompanied by an approximately proportional increase in rw. This is a problem, since an increase in rw means lower signal levels relative to noise and other error sources, resulting in poor frequency and position sensing accuracy. Excitation power levels may be increased to compensate, however this may result in decreased battery life, which is undesirable in a handheld device. In addition, larger or more expensive components may be required to handle the higher power levels.

GB0112332.2 describes one possible solution to this problem, including a two-stage measurement process. A first step comprises an excitation-detection process using small values of EP and DP, for example EP=DP=3, at a fixed excitation frequency Fex chosen to excite a wide range of possible resonator frequencies. This results in a large value for rd, so that the measurement of Pe described above may be unambiguous over a relatively large frequency range. However, the signal levels are low due to a large value of rw, and the results of the measurement may not be acceptable for position or frequency measurement. A second step comprises an excitation-detection process with larger values of EP and DP, for example EP=DP=16. The excitation frequency Fex for this second

step would be set to the approximate resonator frequency determined from the measurement of Pe from the first step. Assuming that there is only a small change in resonator frequency between the first and second steps, the second step will result in both a large signal level and unambiguous frequency indication.

This approach improves frequency and position measurement accuracy, however the improvement may be insufficient for some applications. This disclosure includes options for further improvements.

3 DESCRIPTION

One of the problems with systems described in the previous section was that the range over which pen frequency can be determined unambiguously is limited when phase is used as an indication of pen frequency.

Amplitude information may be used as an alternative. In particular, amplitudes measured at two or more different excitation frequencies will differ, and comparing them yields an indication of the resonant frequency.

The Digital Processing and Signal Generation Unit may calculate an estimate of amplitude from:

$$Ae = \sqrt{SI^2 + SQ^2} \approx A = Ao \cdot e^{-\left(\frac{1-r}{rw}\right)^2}$$

Figure 5 shows how Ae(1) and Ae(2), the results for an excitation-detection process at Fex(1) and Fex(2), typically vary with resonator frequency Fpen. In addition, there is a plot of the functions:

$$APe = \ln\left(\frac{A(1)}{A(2)}\right)$$

$$CPe = a\tan\left(\sqrt{\frac{A(1)}{A(2)}}\right)$$

Both of these are relatively linear with resonator frequency and either may therefore be used to derive an estimate for resonator frequency through re-scaling functions, e.g.:

$$Fpen\_AP\_estimate = Fpen\_AP\_0 + AG \cdot APe$$

$$Fpen\_CP\_estimate = Fpen + \_CP\_0 + CG \cdot CPe$$

where Fpen_AP_0, Fpen_CP_0, AG and CG are constants that depend on the system's configuration. There are other functions that may be used to derive an estimate from input parameters Ae(1) and Ae(2). The general property of these functions is that they are ratiometric and the two variables may be exchanged given appropriate modifications to the offset and gain terms in the relevant re-scaling equation.

Note that this approach results in unambiguous pen frequency detection across a broad range of frequencies, unlike the measurement of phase, provided the system has sufficient accuracy.

Should the accuracy prove insufficient, it is possible to employ more than two excitation frequencies, and to base the resonator frequency estimate on the two measurements that yielded the highest amplitudes. For example, in the case of N excitation frequencies Fex(1), Fex(2)....Fex(N) with calculated amplitudes Ae(1), Ae(2)...Ae(N), the following function may be employed to derive an estimate of Fpen in place of APe above:

$$SFe = \frac{\sum_{n=1}^{n=N} Fex(n) \cdot Ae(N)}{\sum_{n=1}^{n=N} Ae(N)}$$

In the case of 3 excitation frequencies, the following function may be employed in place of APe above, which generates the estimate based on the assumption that amplitude varies approximately parabolic with frequency:

$$MPe = \frac{A(1) - A(3)}{2 \cdot \left(A(1) - 2A(2) + A(3)\right)}$$

The position processing electronics illustrated in figure 1 may typically suffer from offset errors, such that the digital measurement of signal level may be represented approximately as:

$$SI = A \cdot \cos(P) + OI$$
$$SQ = A \cdot \sin(P) + OQ$$

Where OI and OQ are, in general, a function of the particular mixer/integrator channel on a particular device, supply voltage, temperature and time. They may depend on other factors specific to the circuit's implementation, e.g. integration time. These offsets may be sufficiently large to interfere with frequency detection accuracy.

In this case, a calibration step might be employed where the offsets are somehow measured then subtracted from each measurement.

One approach for measuring these offsets is to perform the excitation-detection process omitting the excitation, so that A=0 in the model equation above and SI and SQ yield the offsets directly. A single measurement, or set of averaged measurements, may not yield sufficient accuracy, given variables such as temperature and supply voltage. In this case, measurements may be taken repeatedly, for example before each normal excitation-detection sequence.

An alternative approach is to base each amplitude measurement on two excitation-detection sequences, having the same excitation frequency but with

inverted excitation voltage so that the amplitude term A in the model above is inverted:

$$SIt = SI(+) - SI(-) = A \cdot \cos(P) + O1 - (-A \cdot \cos(P) + O1) = 2A \cdot \cos(P)$$
$$SQt = SQ(+) - SQ(-) = A \cdot \sin(P) + O2 - (-A \cdot \sin(p) + O2) = 2A \cdot \sin(P)$$

The frequency estimating approaches above may be applied to SIt and SQt for each excitation frequency instead of SI and SQ, yielding measurement STAe with improved frequency accuracy. However, the number of excitation-detection processes is doubled, and this may limit the pen position sample rate, and hence the ability of the system to respond to the user pen movements satisfactorily.

The mix signals may be inverted as an alternative to inverting the excitation signal. However, this may not be preferred, since each channel's offset may be a function of the mixer waveform, so that an inverted waveform may yield a slightly different offset and hence imperfect cancellation.

Alternatively, offsets may be eliminated by combining data from different excitation frequencies. Figure 6A shows SI channel measurement results from excitations at 100.5kHz and 106.0kHz, SI(1) and SI(2) respectively, as a function of resonator frequency. These results include approximately 50 units of offset O1. If the two results are subtracted yielding SI(2-1), the offset is cancelled while the signal remains. The wanted signal components of SI(1) and SI(2) are approximately out of phase, such that their subtraction yields a larger wanted signal than either SI(1) or SI(2). This is not a necessary condition, but it is preferred since it results in a measurement with good signal to noise.

Figure 6B shows the corresponding SQ channel measurements and SQ(2-1), the result of their subtraction. The Q channel's offset is cancelled, while maximising wanted signal components, as for the I channels.

Signals SI(2-1) and SQ(2-1) may be approximated by the same amplitude/phase model as signals resulting from a single excitation-detection process above, but note that the model parameters may change. As such, a phase estimate Pe may be calculated and used to estimate resonator frequency as described above. This approach has the advantage of removing errors caused by offsets, and the use of two excitation frequencies broadens the frequency range over which useful signal level is obtained.

Figure 7A shows SAe(2-1), the result of combining SI(2-1) and SQ(2-1) to yield an amplitude figure. Like the measurement STAe above that employs excitation at reversed phase, this amplitude measurement is largely free of error caused by offsets.

A further excitation-detection process may be employed at a third frequency yielding SI(3) and SQ(3). Data from the second and third excitation-detection

processes may be combined in the same way as the first and second described above, yielding SI(3-2), SQ(3-2) and SAe(3-2). Figure 7B shows these functions for the case of a third excitation frequency equal to 111.5kHz. In this case, the offsets are cancelled and wanted signal level is maximised in a similar fashion to frequencies 1 and 2.

The calculated values SAe(2-1) and SAe(3-2) both vary with resonator frequency in a similar fashion to the signals resulting from a single excitation-detection process described above. The same approaches may be employed to estimate frequency from these two amplitude figures, for example the following function varies approximately linearly with resonator frequency:

$$APe(3\_2\_1) = \ln\left(\frac{SAe(2-1)}{SAe(3-2)}\right)$$

The resulting frequency estimate is based on only 3 excitation-detection processes, unlike the approach outlined earlier using excitation signal inversion that required 4. It therefore takes less time to perform. In addition, each excitation is at a different frequency, so that excitation power is not concentrated at particular frequencies by multiple excitations at those frequencies. The consequence is a more even power distribution across frequencies, which is desirable since it minimises the power required to cover a known frequency range with sufficient energy to detect a resonator with sufficient signal level.

This example was for three excitation-detection processes. It is possible to extend the approach to more frequencies using the approaches outlined earlier for dealing with more than 2 functions of amplitude, and by combining measured data to eliminate offsets while maximising wanted signal levels.

The resonator frequency estimates determined from amplitude information described above may be improved by combining them with frequency estimates based on phase information. An estimate based on amplitude covers a broad frequency range unambiguously, but it typically relatively inaccurate. An estimate based on phase is accurate, but suffers from ambiguities as described earlier.

In the case of three excitation frequencies, the pairs SI(2-1) and SQ(2-1), and SI(3-2) and SQ(3-2) may be combined to obtain two phase estimate by calculating Pe as illustrated above for each pair. Alternatively they may be combined to yield a single phase estimate as illustrated in figure 8. This latter approach has the advantage of yielding a single function SPe(3_2_1) and requiring only a single inverse tangent calculation.

In this case, phase information SPe(3_2_1) may be combined with amplitude information APe(3_2_1) using the following calculation, in the case where coupling factors between pen and sensor coil may become inverted:

$$CPe = SPe(3\_2\_1) + \pi \cdot floor\left(\frac{GA \cdot APe(3\_2\_1) - APo - SPe(3\_2\_1)}{\pi}\right)$$

GA is a constant such that the rate of change of APe(3_2_1) multiplied by GA with resonator frequency is nominally equal to the rate of change of SPe(3_2_1) with frequency. APo is a constant that is chosen so that the result of the floor() calculation nominally increments at the same resonator frequency as the SPe(3_2_1) calculation wraps, to ensure maximum margin for errors.

In the case where it is known that the coupling factor between pen and sensor coil does not invert, the two instances of π above may each be replaced by 2π.

CPe may be scaled and offset as described earlier to yield accurate resonator frequency without phase wrapping in a broad frequency range of interest.

Just as more than 2 functions of amplitude may be combined to yield an approximate resonator frequency, I and Q information may be combined from measurements at multiple frequencies. For example, a single pair of I and Q values could be chosen from several pairs at different frequencies by selecting the values that yield the highest amplitude figure. Alternatively, data from several frequencies could be combined by calculating a weighted average, where amplitude is used for the weighting function.

The offsets OI and OQ described above may not remain constant across excitation frequencies, for example if they depend on integration time. In this case the approach above may not result in ideal cancellation of offsets. This can be solved by modifying the calculation of SI(2-1):

$$SIc(2-1) = K2 \cdot SI(2) - SI(1)$$

K" is selected for minimum overall error. In the case of offsets that are exactly proportional to integration time, K2 may be determined from:

$$K2 = \frac{\text{integration time } frequency\ 1}{\text{integration time } frequency\ 2}.$$

And similarly for other channels and frequencies. This approach may be undesirable since it requires a relatively lengthy multiplication step. As an alternative, a fixed offset may be added:

$$SId(2-1) = SI(2) - SI(1) + OI(2-1)$$

Where

$$OI(2-1) = OI(2) - OI(1) \approx OI(1) \cdot \left(\frac{\text{integration time frequency } 2}{\text{integration time frequency } 1} - 1 \right)$$

In this case the majority of the offset is removed by the subtraction of SI(1) from SI(2) because the integration times are typically similar. This is advantageous because it means variations in offset over temperature, supply voltage and so on are largely eliminated.

Figures 6 to 8 above are for fixed excitation and detection phases; they do not change across the three excitation frequencies. It is possible to vary the phase of either the excitation or detection signals (in phase and quadrature out) or both for each excitation-detection process. In this case, the difference between excitation frequencies is preferably modified so that the offset cancellation calculations also yield maximum signal level. The frequency difference for optimum signal level may be increased, by suitable choice of excitation and/or detection phase. This approach may be employed to extend the range of excitation frequencies, therefore covering a greater range of possible resonator frequencies.

The approach described above uses a single coil to detect resonator frequency. This may be appropriate for a pen-sensing system that includes a sensor coil that is coupled to the resonator with a sufficiently large coupling factor across all pen positions of interest. For example, the outer coil described in WO 00/33244, which is also used for excitation. However, none of the loop coils of US 4,878,553 are always coupled to the pen with sufficiently large coupling factor across all pen positions of interest. In the case of WO 00/33244, if only the four inner coils SX, CX, SY and CY are connected to processing electronics, none of these is guaranteed to have sufficiently large coupling factor across all pen positions of interest. Although the coils cover the entire sensing area, they have regions of zero coupling factor by design. In these cases, it is necessary to combine the information from more than one sensor coil channel pair.

For pen positions of interest, at least one of the SX and CX coils of WO 00/33244 is coupled to the resonator in the pen. Sensor coil SX would be connected to a pair of input channels driven with in-phase and quadrature mix signals, yielding integration results SXI and SXQ respectively. Sensor coil CX would be connected to a pair of input channels driven with in-phase and quadrature mix signals, yielding integration results CXI and CXQ respectively. The following description illustrates how these signals may be combined to yield signals SI and SQ, suitable for processing according to the description above.

For the purposes of amplitude interpolation, for example for calculating each of SAe(2-1) and SAe(3-2), SXI, SXQ, CXI and CXQ signals may be combined by extending the equation for Ae as follows:

$$SCXAe = \sqrt{SXI^2 + SXQ^2 + CXI^2 + CXQ^2}$$

Other combinations may be used, for example by summing the moduli:

$$SCXAe' = \left|\sqrt{SXI^2 + SXQ^2}\right| + \left|\sqrt{CXI^2 + CXQ^2}\right|$$

For the purposes of phase measurement, for example determining SPe(3_2_1), it is possible to determine phase from whichever of the SX or CX data is most reliable, for example by determining which has greater signal strength, by some measure. For example:

$$MSX = \sqrt{SXI^2 + SXQ^2}$$
$$MCX = \sqrt{CXI^2 + CXQ^2}$$

These measures of signal strength require relatively complicated calculation and may be replaced by the following without significant loss in accuracy, since the ratio of I to Q signal strengths in the two coils are approximately equal:

$$MSX' = |SXI| + |SXQ|$$
$$MCX' = |CXI| + |CXQ|$$

Information from the channel with maximum signal strength may then be selected:

$$MXI = SXI \text{ IF } MSX' \geq MCX' \text{ ELSE } CXI$$
$$MXQ = SXQ \text{ IF } MSX' \geq MCX' \text{ ELSE } CXQ$$

It is possible to use MXI and MXQ from the equation above for determining electrical phase, by substituting them in place of SI and SQ into the equation for Pe defined earlier. This approach is simple, but information from the coil with the lower measure of signal strength is lost, particularly at resonator positions where MSX' and MCX' are approximately equal. If signal strength was marginal in the first place, this may mean the difference between correct and incorrect frequency detection. As an alternative, information from the coil with the lower measure of signal strength may be included in the phase measurement by combining it with MXI and MXQ above. It is helpful to define LXI and LXQ as follows:

$$LXI = CXI \text{ IF } MSX' \geq MCX' \text{ ELSE } SXI$$
$$LXQ = CXQ \text{ IF } MSX' \geq MCX' \text{ ELSE } SXQ$$

It is not appropriate to combine this information by directly adding LXI to MXI and LXQ to MXQ, because the SX and CX coil to resonator coupling factors may

have opposite sign. The relative sign may be determined by comparing phase angles as follows:

$$RSIGNSC = 1 \ IF \ -\frac{\pi}{2} \leq a\tan 2(MXI, MXQ) - a\tan 2(LXI, LXQ) < \frac{\pi}{2} \ ELSE \ -1$$

RSIGNSC should return 1 if SX and CX coupling factors are the same sign and -1 if not. The following is equivalent, and may prove faster to implement:

$$RSIGNSC = SIGN[MXI \cdot LXI + MXQ \cdot LXQ]$$

Alternatively, for relatively little impact on system accuracy, the following may be used for simplicity and speed of calculation:

$$RSIGNSC' = 1 \ IF \ (|MXI| + |MXQ|) \geq (|MXI - LXI| + |MXQ - LXQ|) \ ELSE \ -1$$

The pairs of data may now be combined as follows:

$$SI = MXI + RSIGNSC' \cdot LXI$$
$$SQ = MXQ + RSIGNSC' \cdot LXQ$$

The process for combining SXI, SXQ, CXI and CXQ data to yield combined data SI and SQ may be applied to any set of I and Q measurements from the two coils. For example, to the difference between measurements at different excitation frequencies SXI(3_2_1), SXQ(3_2_1), CXI(3_2_1) and CXQ(3_2_1), where the suffix (3_2_1) implies the same differencing procedures defined above for SI(3_2_1) and SQ(3_2_1). The resulting combined data SI(3_2_1) and SQ(3_2_1) may then be used to determine an improved estimate of phase, and therefore frequency.

This combination process may be employed for data from any two coils. If the signal strength in either or both of those coils is sufficient then a reliable phase estimate may be obtained, even if one of the coils has zero signal strength.

The same procedure maybe adopted when there are many coils, for example in the case of US 4,878,553. The coil with maximum signal strength may be chosen as above, and additional coil signals would be either added or subtracted depending on its relative phase as determined using either of the tests for phase above for determining RSIGNSC. This approach may also be adopted for all four sensor coils described in WO 00/33244, and may improve overall signal strength and hence system robustness for a given power level.

In order to calculate position from a sensor board designed according to WO 00/33244 or US 4,878,553 operating with the electronic processing approach described in this document, a number representing coupling factor between the

resonator and each sensor coil is required. The approach above yields I and Q numbers for each channel. It is possible to choose the signal with greatest magnitude, but this does not necessarily result in best performance. It may not yield minimum noise and electronic errors, because information from the other channels would be discarded.

In addition, eddy currents flowing in any conductive sensor board screening may generate undesirable signals. These signals typically have a constant phase relationship to the wanted resonator signal. Ideally the impact of these undesirable signals should be relatively constant and not a function of pen frequency and detection phase, otherwise they may be difficult to eliminate by conventional means.

In order to achieve this, it is preferable to synthesise a new signal from each coil's I and Q measurement which is equivalent to the length of the (I,Q) vector resolved at a phase angle PR. PR may be selected for optimum system performance. PR may equal CPe as calculated above; in the case that maximum signal strength is ideal. Alternatively, it may be equal to CPe plus some constant, where that constant is chosen to minimise the effect of some known error source such as eddy currents.

As an example, the following equation may be employed to determine the length of the (SXI(3_2_1), SXQ(3_2_1)) vector along a line at angle CPe to the origin:

$$SX(3\_2\_1) = SXI(3\_2\_1) \cdot \cos(CPe) + SXQ(3\_2\_1) \cdot \sin(CPe)$$

This is equivalent to a coordinate rotation of CPe about the origin. Other translations may be employed. For example, the coordinate rotation may be limited to a small number of possible angles, say 8 or 16 per 2π radians. The closest rotation angle to CPe would be chosen. The advantage of this approach is to simplify the calculation above, for example by restricting multiplies to ones that are straightforward for a microcontroller with limited instruction set and speed. For example, coordinate rotation by 0.896 radians can be achieved by substituting cos(CPe)=1.25 and sin(CPe)=1 in the equation above. This results in an amplitude error, but this may be insignificant provided the same rotation process is applied to all coils and subsequent calculations are ratiometric. Multiplication by 1 is trivial, and by 1.25 is relatively straightforward since it requires a shift by 2 bits in binary representation followed by one addition.

Note that by resolving at angle CPe in this way SX(3_2_1) does not flip sign with frequency, unlike SXI(3_2_1) and SXQ(3_2_1). Note also that if CPe were replaced by SPe, the phase estimate based on phase alone and excluding amplitude, SX(3_2_1) would flip sign due to phase wrapping at $\pm\pi/2$ radians.

To continue the calculation of position from a sensor board designed according to WO 00/33244 or US 4,878,553, the resolving process described above would be

applied to each set of coil signals of interest. In the case of WO 00/33244, X and Y position estimates may then be calculated from:

$$X = X0 + X1 \cdot a\tan 2(CX(3\_2\_1), SX(3\_2\_1))$$
$$Y = Y0 + Y1 \cdot a\tan 2(CY(3\_2\_1), SY(3\_2\_1))$$

In the case of US 4,878,553, position would be calculated using the appropriate interpolation functions based on the resolved amplitude figures for each coil.

The position calculation approach above has been described with respect to SXI(3_2_1), SXQ(3_2_1) and equivalent measurements for the other coils in the system. The approach is also appropriate for any measurement of coil signal level described above, for example measurements based on the difference between I and Q data from excitation-detection processes 1 and 2.

Once resonator frequency has been estimated using the approaches above, it is possible to perform an additional excitation-detection process with Fex equal to this frequency estimate, in order to further improve the frequency estimate and to accurately determine resonator position. A large number of excitation and detection pulses may be chosen for this second step, yielding high signal levels without the risk of phase wrapping, except in the case of significant resonator frequency shift between the two parts of the process.

If this process yields a phase PNe, the frequency estimate may be corrected as follows:

$$RFE = Fex - GN \cdot PNe + Fd$$

Where RFE is the final resonator frequency estimate, Fex is the final excitation frequency, GN is the rate of change of PNe with resonator frequency and Fd is an offset to compensate for system phase and frequency offsets and errors. In this case, position may be calculated from the final detected signals using the approaches above.

The approaches outlined above may fail if resonator frequency changes significantly between two or more of the excitation-detection processes. This may be acceptable providing false frequency and position data is not reported to the host system. It may therefore be desirable for the pen sensing system to perform tests to verify data is correct. In the case of a two step process, one possible test is to verify that position and frequency calculations for each step are consistent. As the position and frequency measurements from the first step are relatively coarse, and the resonator position may change during the measurement too, reasonable error must be allowed for.

A disadvantage of the two step approach outlined above is that several excitation-detection processes are required. The first step is relatively wasteful of

both time and power in the case of a pen sensing system, because much of the time the pen frequency does not change between positions samples, especially at high sample rates, for example 100Hz. The first step could be omitted once frequency has been reliably detected, and the resonator frequency estimate could be based on the second step alone. The resonator frequency estimate RFE could still be updated each time using the equation for RFE above, enabling the system to track subsequent pen frequency changes.

However, if the pen frequency changed very rapidly or a low sample frequency were required for power saving such that the time interval between samples enabled significant frequency change, the phase could wrap, leading to frequency ambiguity as described earlier. This fault could be detected by observing the subsequent change in position indication or amplitude, but a system based on these tests might not be sufficiently reliable. For example, if the pen approaches the writing surface rapidly and is sampled once at some distance unclicked then a second time clicked so that the frequency changes significantly, Ao will have increased significantly. In combination with the amplitude reduction due to the change in frequency modelled above, the reported amplitude Ae might not change significantly. The phase may have wrapped twice such that the phase and position indication remains the same. The system may therefore be unable to tell the difference between the first and second data.

This ambiguity may be resolved by arranging for the processing electronics to have channels with two or more different rates of change of phase with frequency, or equivalently two different values of rd. For example, the mixers 69 and integrators 71 of figure 1 could be repeated, with the second set driven by a second RX gate signal with a different duration. Each coil would now be connected to four input channels, comprising I and Q at two different RX gate durations, here labelled A and B.

Figure 9A illustrates channel outputs for the case EP = 16, and DP = 16 for A channels and DP = 24 for B channels. Figure 9B illustrates the calculated amplitudes and phases for the A and B channel pairs. Pe(B-A) is defined as follows:

$$PWe(B-A) = Pe(B) - Pe(A)$$

$$Pe(B-A) = PWe(B-A) +$$

$$\pi \cdot \left[\left(1 \; IF \;\; PWe(B-A) < \frac{-\pi}{2} \; ELSE \; 0 \right) - \left(1 \; IF \;\; PWe(B-A) \geq \frac{\pi}{2} \; ELSE \; 0 \right) \right]$$

Note that Pe(B-A) is unambiguous over a much greater frequency range than either Pe(A) or Pe(B). However, a frequency estimate based on Pe(B-A) is relatively inaccurate. The accuracy may be improved by combining Pe(B-A) with Pe(B) or Pe(A), or some function of the two, whichever yields the best accuracy. That combination may be performed in the same way as SPe(3_2_1) is combined

with APe(3_2_1) above to yield CPe(3_2_1) above. Data representative of individual coil coupling factors can be obtained in the same way as SXI(3_2_1) is combined with SXQ(3_2_1) to yield SX(3_2_1) above, using the combined phase number.

This approach may therefore yield position and frequency data for resonator frequencies substantially beyond the wrapping points of Pe(A) or Pe(B) individually. If the frequency difference is sufficient, Pe(B-A) may wrap and there will be an ambiguity. However, at this point amplitude is very substantially lower than before, and a system with appropriate amplitude threshold will easily detect that the data is unreliable. In this case, the original two step approach could be used for the next position sample.

This last approach doubles the number of electronics detection channels, and is therefore costly and power hungry. The system described in WO 00/33244 includes two coils for measuring in the X direction, SX and CX, and two for the Y direction, SY and CY. As described above, the SX and CX coils may be sufficient to determine phase, with the resonator at any point of interest over the sensor board. The same goes for the SY and CY coils as a pair. It is therefore possible to drive the X and Y coil channels with an RX gate signal of different duration, and to determine Pe(A) from the X coils and Pe(B) from the Y coils. This way, the number of electronic processing channels is minimised. The same approach may be employed for the sensor board layout of US 4,878,553, where the X and Y coils form two independent sets, each of which sets could be used individually to determine phase.

There are other approaches for achieving different model values rd for different electronic processing channels, other than a difference in the RX gate time. For example, the time between the end of the excitation cycles and beginning of RX gate may be modified, or the channels' gains may be modulated differently as a function of time during the integration process.

Another option is to employ two sets of mixer signals, in-phase out and quadrature out, for the two sets of channels, where each set is at a different frequency. Approximate resonator frequency may be calculated from the relative amplitude of signals from each set of channels, and this information may be combined with phase information from one or both sets to determine resonator frequency more accurately and without phase wrapping.

Some applications require a pen-sensing mode that is used to wake up the host system from standby when an approaching pen is observed. The resonator detection strategies described above are optimised for accurate position and frequency detection, but generally require more than one excitation-detection sequence and are therefore relatively power hungry. As an alternative, for very low power resonator sensing, a single excitation detection sequence may be employed with sufficiently broad frequency range to detect any resonator of interest. The effects of offsets may be eliminated by comparing successive data,

rather than observing when some combination has exceeded a threshold. If there is a significant difference between the current data and historical data, for example the immediately previous data, a flag may be set to indicate that the pen has appeared. This approach does not yield an accurate resonator frequency or position estimate, but may be sufficient to wake up the host system and initiate an alternative resonator sensing mode.

The table below lists some of the methods described above for detecting a resonator with processing hardware similar to figure 1, together with the features of each type of detection:

Method	Advantages	Disadvantages	Applications
Single excitation-detection process	Low power	No frequency or position	"Pen coming into range" detection
3 frequency excitation-detection process (Figures 6,7,8)	Unambiguous frequency detection	May yield poor frequency and position data in case of low signal level	Determining pen frequency for the first time
2 phase slope (Figure 9)	Large signal levels, definite indication of phase wrap and hence unreliable frequency and position	Low signal strength if resonator frequency very different from excitation frequency	Improving "3 frequency" estimate of frequency and phase, regular use once frequency lock is established

The pen sensing system may be designed to operate in any set of appropriate operating modes, for example those listed in the table above, and to choose the most appropriate given the state of the host system and the system's knowledge of the current resonator frequency and position. In this way, the pen sensing system may be optimised for overall low power consumption while maintaining rapid and accurate response when required.

For example, the single excitation-detection process may be operated twice a second until a significant change in signal levels is observed. At this point, the 3 frequency excitation-detection process may be employed to determine resonator frequency and position. This process may be repeated if the resonator signal level is below some first threshold level, say at 5 samples per second for faster response.

If the signal becomes lower than some second threshold lower than the first, selected to test for resonator absence, the system may revert to the single excitation-detection process.

A third threshold may be selected, above which the resonator's frequency and position will be determined accurately. If the signal exceeds this third threshold, the system proceeds to the 2 phase slope approach. This mode of operation may be repeated at, say, 100Hz until either the signal level has reduced below some fourth threshold level or the system loses track of resonator frequency, for example due to a rapid change in frequency. In this case, the system may revert to the 3 frequency excitation-detection process.

The host system may only require position data when the pen is clicked. If the frequency estimate generated by the 3 frequency excitation-detection process is within a first set of frequency limits, determined to test for pen unclicked and including an allowance for possible frequency error of the 3 frequency excitation-detection process, it is not necessary to waste power by proceeding to the 2 phase slope mode in order to refine the frequency and position estimate, since it is already clear that the pen is unclicked.

Information generated by one detection method may be used to improve the performance of the following detection. For example, a measurement of signal level may be used to optimise the sensitivity of the following detection process. That optimisation may take the form of a modification to the excitation power level, integrator gain and the number of excitation and detection pulses, such that the ratios of signal to noise and signal to errors are maintained at an acceptable level. For example, if the amplitude detected by a particular process were below a certain threshold, the following detection process could be performed with an increased power level.

The description in this document so far has been for an excitation process immediately followed by detection. The approaches above may also be appropriate where some or all of the detection process occurs during the excitation process, although this is not preferred due to potential coupling between excitation and sensor coils, which may induce significant errors.

The description in this document so far has been for two mixer phases 90° apart. It is also applicable to a greater number of phases, and where those phases are not necessarily 90° apart. The calculation of amplitude described in this document could be modified to include information from multiple phases, for example by determining the square root of the sum of the squared signal level from each phase, to replace the sum of the squares of I and Q data. The calculation of phase could be updated to resolve any appropriate phase information into two orthogonal phases, followed by an inverse tangent on these two numbers in the normal way. An example of this approach is described in WO 99/18653.

The electronics shown in figure 1 includes multiple instances of mixer plus integrator, to enable multiple data to be measured from a single excitation. This

is an advantage where the excitation process consumes significant power, and where the time for performing resonator measurements is limited. As an alternative, two or more sensor coils may be multiplexed into a single mixer/integrator, and data for each coil can be measured using separate excitations until sufficient data has been accumulated. Similarly, a single channel may detect at both I and Q phases by performing two separate excitation processes, with the mixer control being switched between in-phase and quadrature between the two.

The integrator may be replaced by some other circuit that combines information from the mixer over more than, say, two cycles of a typical resonator frequency.

As an alternative to synchronous detection using a mixer, the processing electronics may detect peak resonator amplitude. If a phase measurement were also required, this may be obtained by measuring the zero crossing point of the resonator signal.

The mixing and integration process may be performed in digital electronics, with the raw sensor coil signals being fed directly into an analog to digital converter.

It is possible to apply the approaches outlined in this document to systems where the phase and frequency of the signals driving the mixers do not match the phase and frequency used for excitation. There may be advantages to alternative schemes, and there is a description above of a system employing two different detection frequencies simultaneously.

The systems described in WO 00/33244, GB0112332.2 and above are typically for a pen whose frequency changes with nib pressure, such that there is a frequency above which the pen is assumed clicked and below which unclicked. Once the system has determined frequency, it performs this comparison and outputs a flag whose state indicates whether the pen is clicked or unclicked.

This system may also be used with a pen whose frequency changes smoothly with frequency, and such a system could output a number representing resonator frequency and therefore indicative of tip pressure.

A pen may be constructed whose resonator frequency depends on more than one variable, for example both tip pressure and the state of some other switch mounted on the pen, for example as manufactured by Wacom. If it is possible to define a function of frequency that determines the state of each of these variables, that function may be performed by the Digital Processing and Signal Generation Unit and the results reported to the host system.

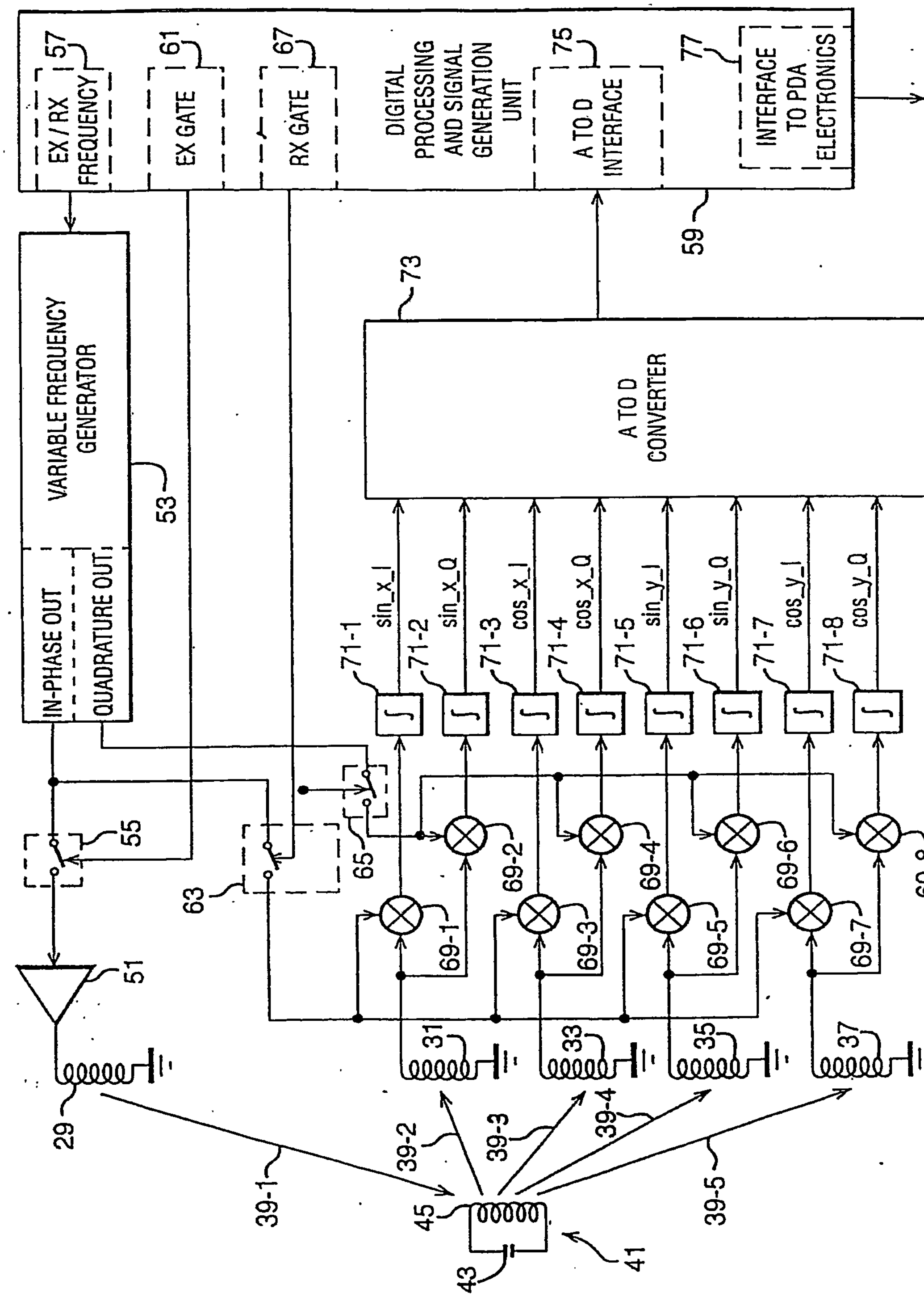
A pen may include two resonators, one at the writing tip and one at an eraser end, at different frequencies, for example as manufactured by Wacom. The systems and approaches described in this document may be used to determine the frequency of the resonator in proximity with the sensor board, and therefore

whether the user requires writing or erasing action, depending on the resonator frequency detected.

The resonator frequency detection techniques described in this document are appropriate for applications other than pen sensing. For example, the system could be used in conjunction with the sensor and resonator designs presented in WO 99/18653 for determining the position and speed of a motor's rotor. If the resonator were designed such that its resonant frequency were determined by some variable whose measurement was required, and where the most convenient mounting point for an appropriate transducer were the moving rotor, the resonator detection system above could be used to determine the value of the parameter of interest, without the need for slip rings for connection to the transducer.

For example, a transducer could be constructed including an inductance or capacitance that depended on rotor torque. In combination with a suitable resonating capacitance and/or inductance and a resonator coil arrangement suitable for position sensing, the Digital Processing and Signal Generation Unit may report both position and motor torque to the motor controller. This approach may yield improvements to product cost by minimising components such as slip rings and by centralising electronic processing and signalling into a single device.

Figure 1



EX / RX FREQUENCY
57
EX GATE
61
RX GATE
67
DIGITAL PROCESSING AND SIGNAL GENERATION UNIT
A TO D INTERFACE
75
INTERFACE TO PDA ELECTRONICS
77
59
VARIABLE FREQUENCY GENERATOR
53
IN-PHASE OUT
QUADRATURE OUT
73
A TO D CONVERTER
sin_x_I
sin_x_Q
cos_x_I
cos_x_Q
sin_y_I
sin_y_Q
cos_y_I
cos_y_Q
71-1
71-2
71-3
71-4
71-5
71-6
71-7
71-8
55
65
63
69-1
69-2
69-3
69-4
69-5
69-6
69-7
69-8
51
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33
35
37
39-1
39-2
39-3
39-4
39-5
45
43
41

Figure 2

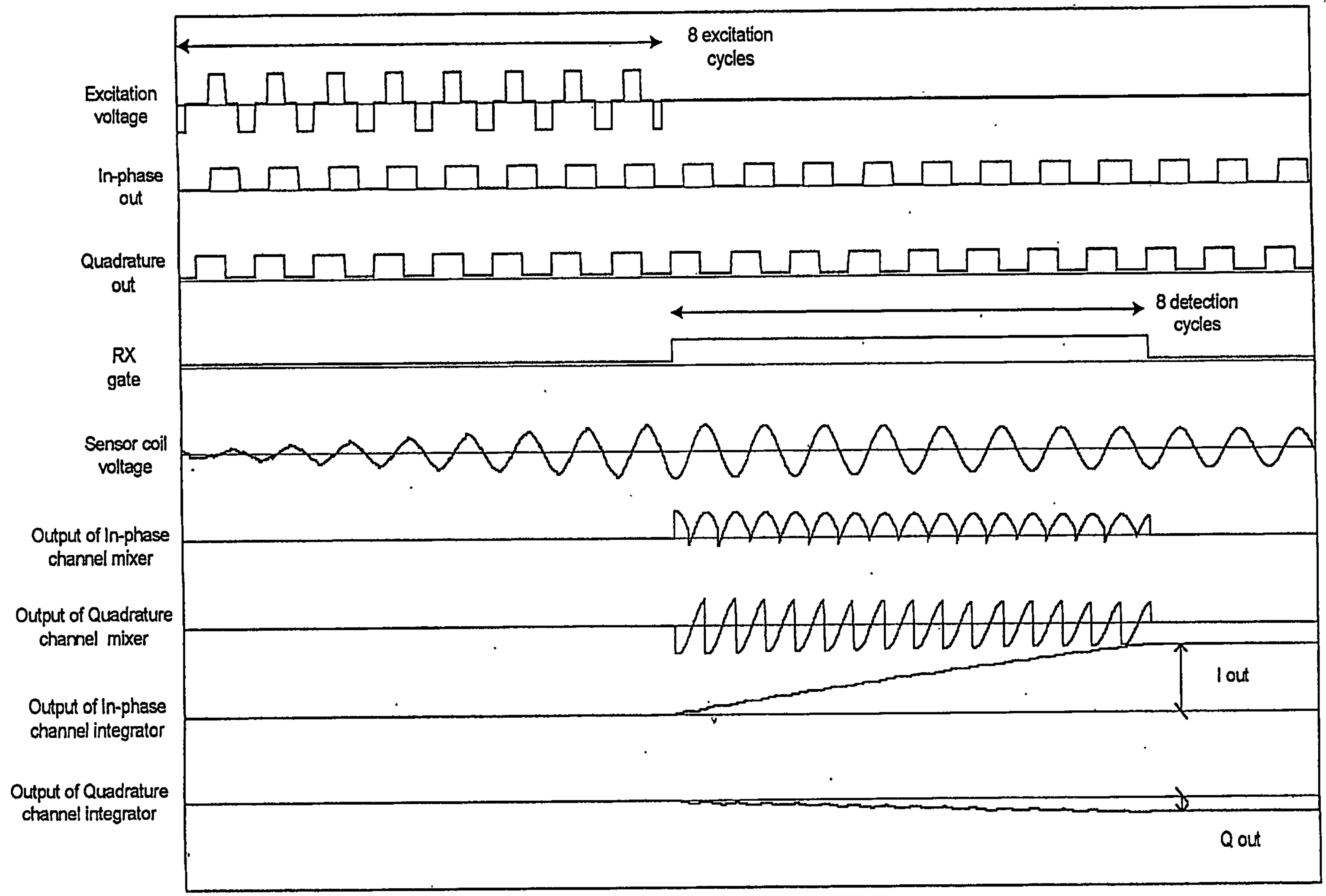
8 excitation cycles
Excitation voltage
In-phase out
Quadrature out
8 detection cycles
RX gate
Sensor coil voltage
Output of In-phase channel mixer
Output of Quadrature channel mixer
I out
Output of In-phase channel integrator
Output of Quadrature channel integrator
Q out

Figure 3A

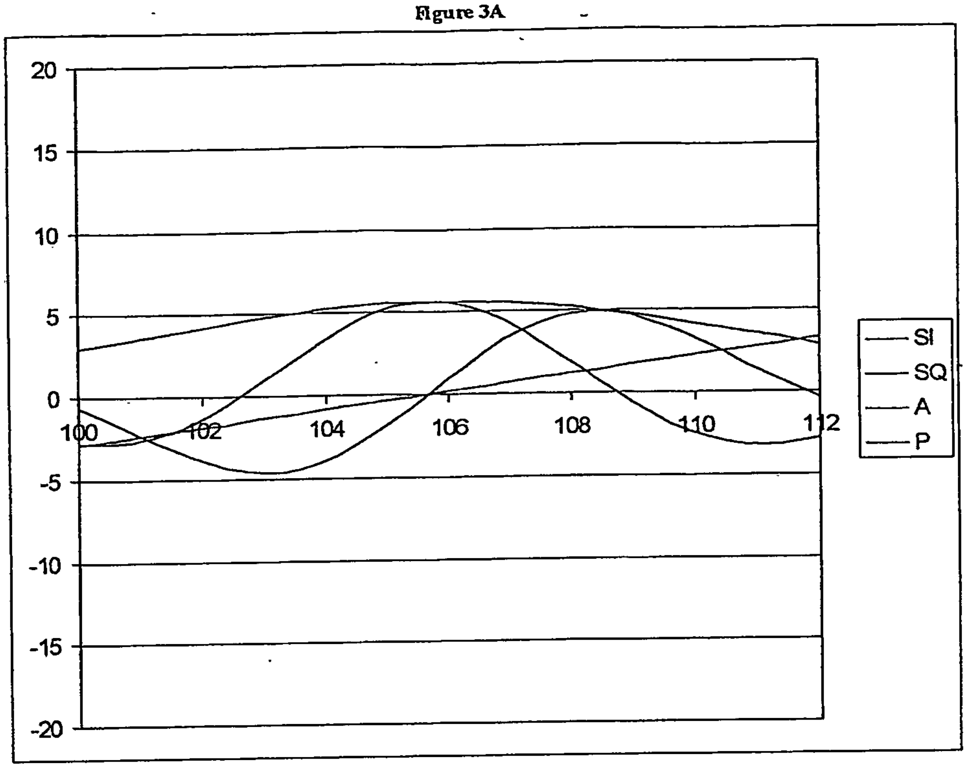

Figure 3B

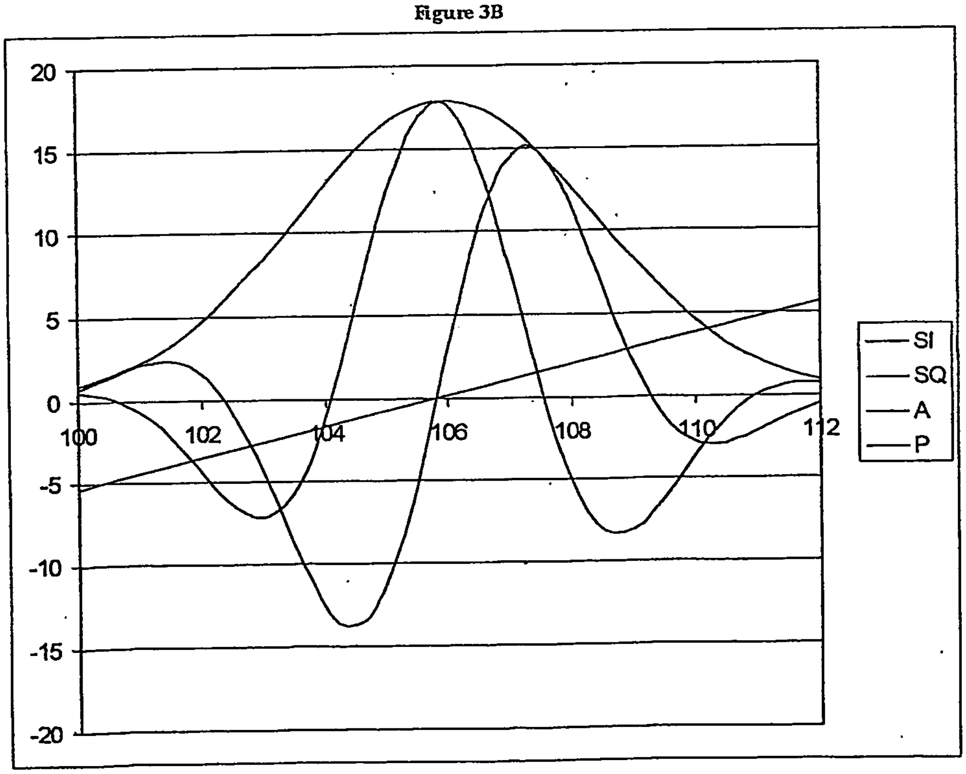

Figure 4

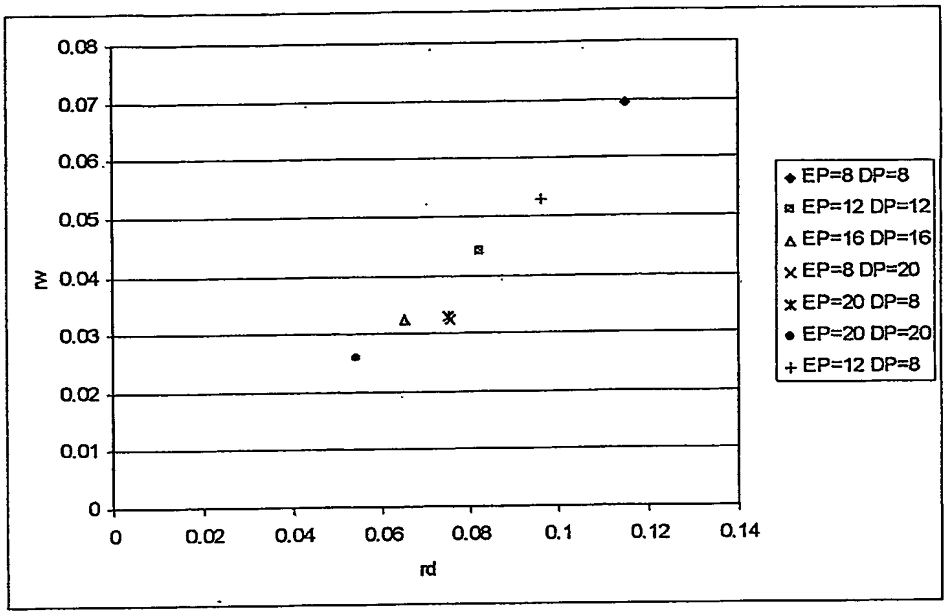

Figure 5

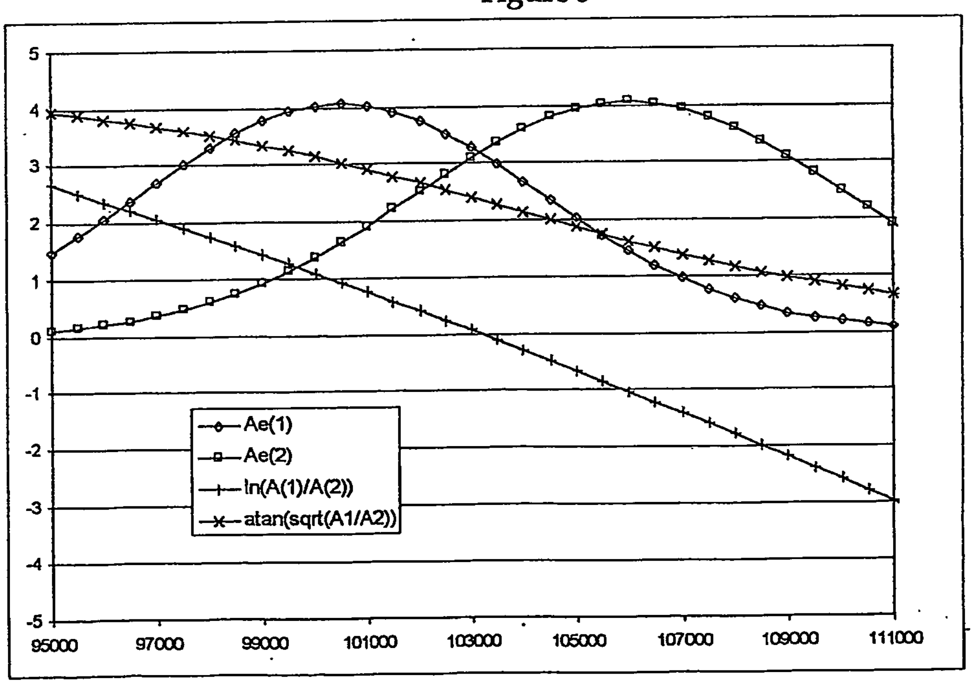

Figure 6A

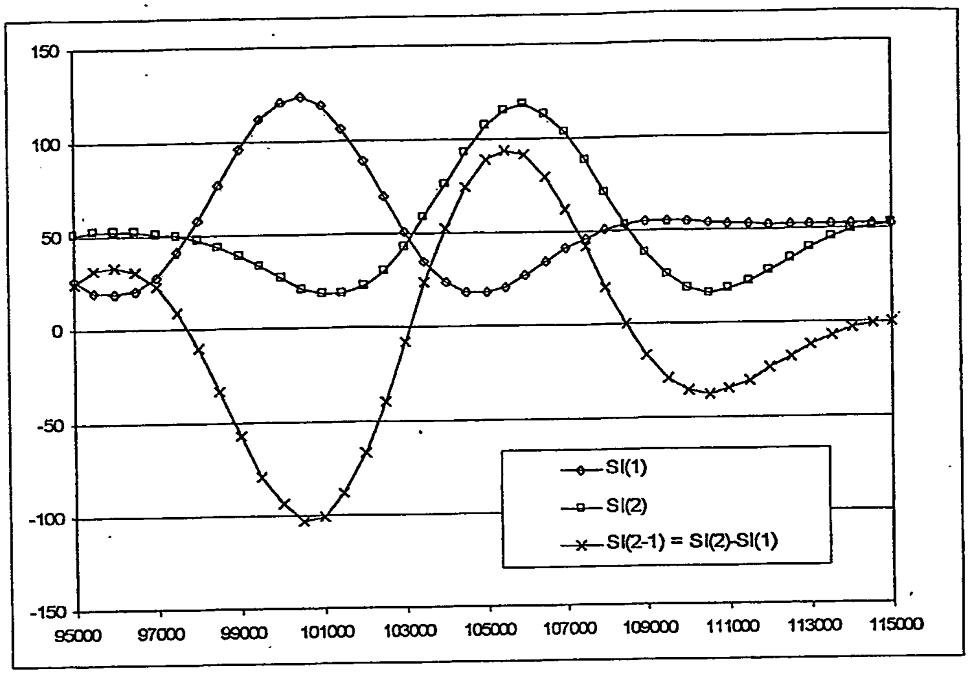

Figure 6B

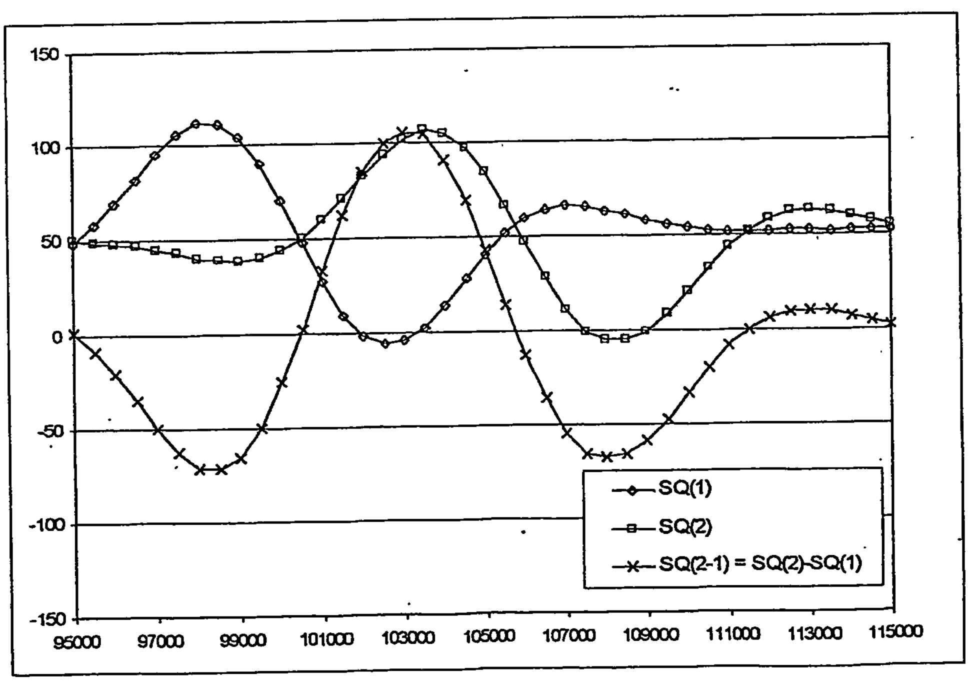

Figure 7A

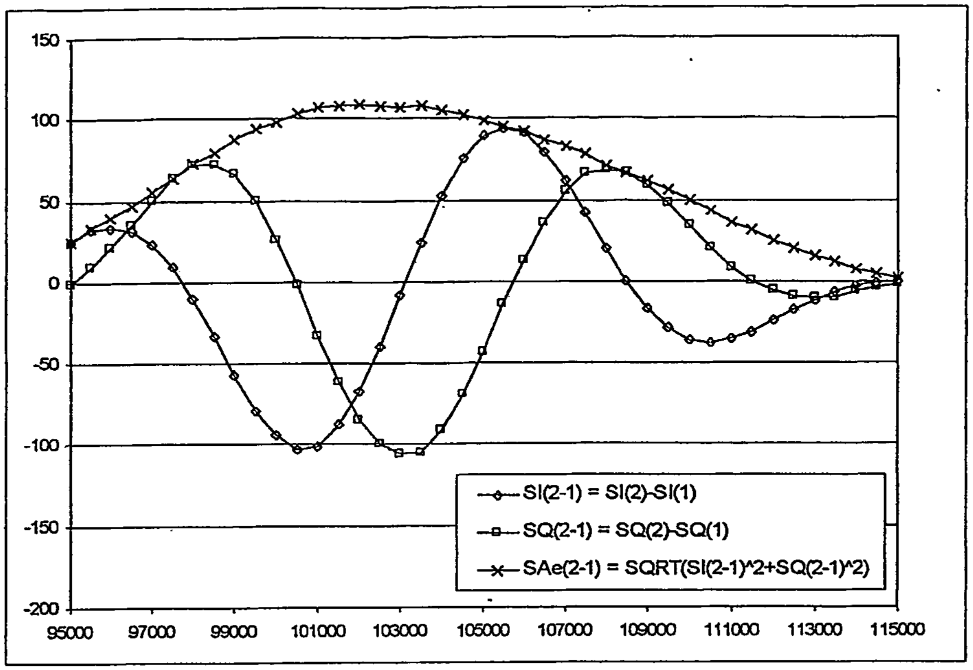

Figure 7B

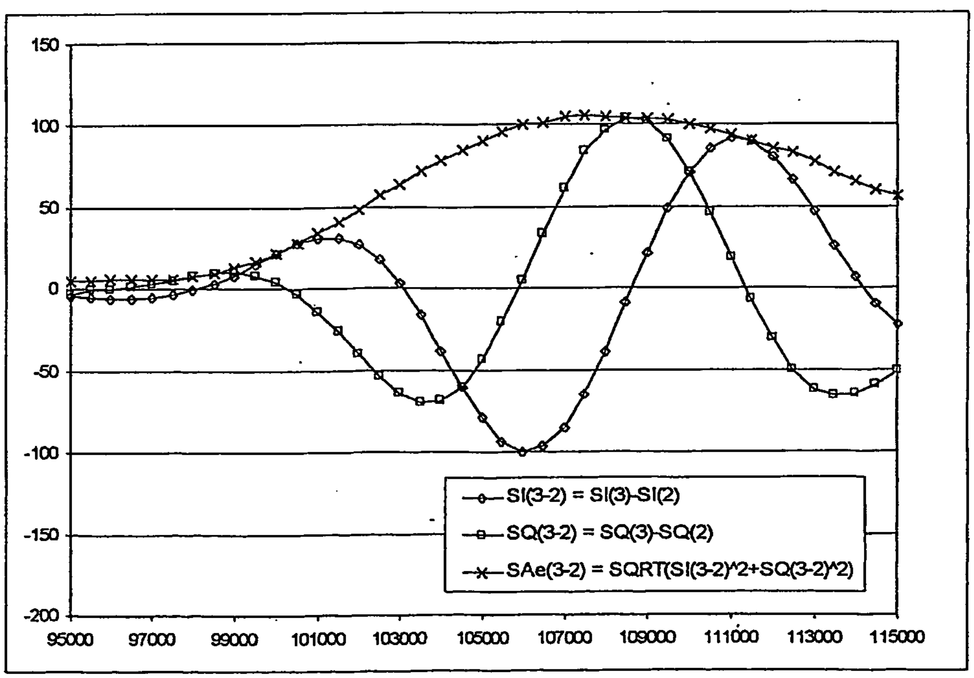

Figure 8

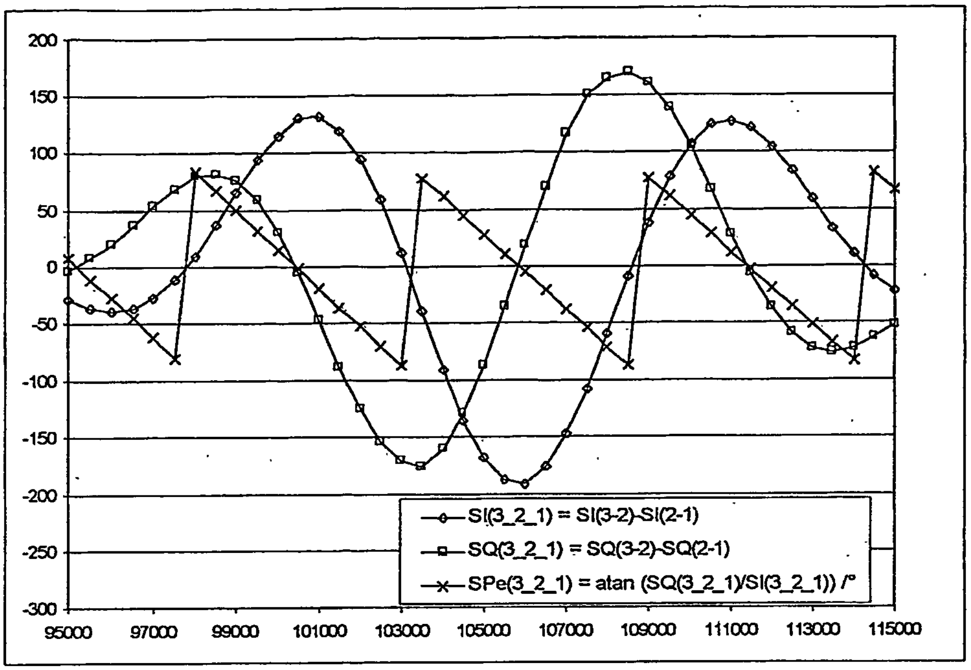
200
150
100
50
0
-50
-100
-150
-200
-250
-300
95000
97000
99000
101000
103000
105000
107000
109000
111000
113000
115000
SI(3_2_1) = SI(3-2)-SI(2-1)
SQ(3_2_1) = SQ(3-2)-SQ(2-1)
SPe(3_2_1) = atan (SQ(3_2_1)/SI(3_2_1)) /°

Figure 9A

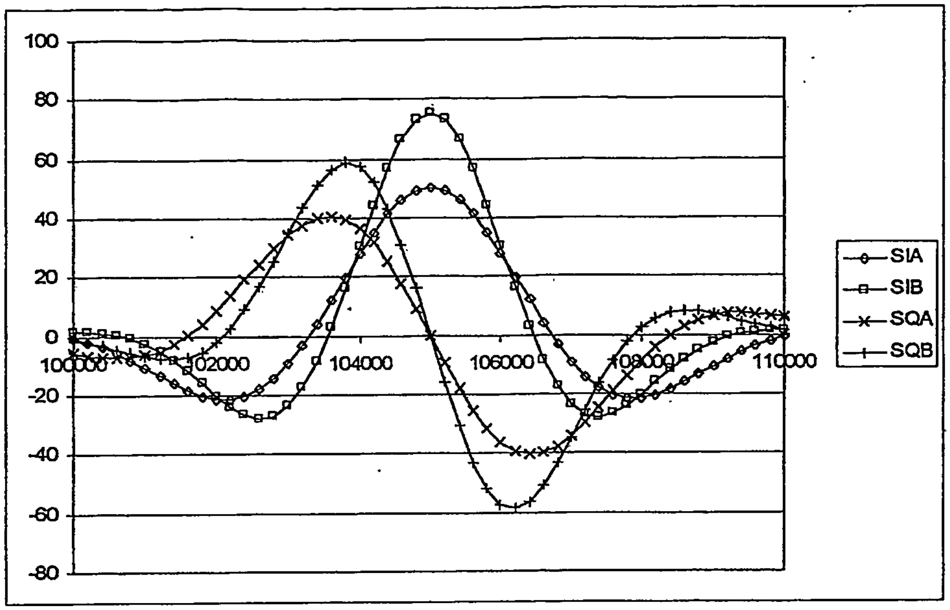

Figure 9B